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Novel fast **GPS**/GLONASS code-acquisition technique us update rate **FFT**

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Abstract:
A novel 'differential' decoding technique is proposed which enables pre-averagi of postintegration for a substantially low update rate '**FFT**-IFT' correlation in spread-spectrum (navigational) receivers. N-channel code acquisition can be p to monitor the time dispersion with **FFT** time left to analyse frequency dispersio reflective areas (e.g. an urban environment).

Index Terms:
differential decoding technique; frequency dispersion analysis; fast code acquis technique; **GPS**; GLONASS; fast code acquisition; code-acquisition technique; rate **FFT**; pre-averaging; N-channel code acquisition; time dispersion; highly r areas; urban environment; fast Fourier transforms; radionavigation; satellite r systems

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assuming that after $i-1$ transmissions the first negatively acknowledged codeword is in position $b_{i-1}+1$, is given by

$$q(b_i/b_{i-1}) = P_D(1-P_D)^{b_i-b_{i-1}} \tag{1}$$

where P_D is the probability of detecting in error a codeword belonging to code C. The probability $P_s(i)$ that after the ith transmission ($i \geq 1$) all the codewords of the jth group have been recovered can be expressed as

$$P_s(i) = \sum_{b_1=0}^{M-1}\sum_{b_2=b_1}^{M-1}\cdots\sum_{b_{i-1}=b_{i-2}}^{M-1} P_D(1-P_D)^{b_1} \times P_D(1-P_D)^{b_2-b_1}\cdots P_D(1-P_D)^{b_{i-1}-b_{i-2}}(1-P_D)^{M-b_{i-1}} \tag{2}$$

Therefore, it is

$$P_s(i) = \binom{M+i-2}{i-1} P_D^{i-1}(1-P_D)^M \tag{3}$$

We now evaluate the number n_i of time slots required to transmit the jth codeword group. During the ith transmission ($i \geq 1$) of the jth group, the number n_i of time slots in the ith transmission of the jth group is

$$n_i = N + b_i - b_{i-1} \tag{4}$$

If after i transmissions, the jth group is recovered (i.e. $b_i = M$), it can be easily shown that

$$n_i = i(N-1) + M \tag{5}$$

The mean number n_s of time slots required to transmit the jth group is

$$n_s = \sum_{i=1}^{\infty} n_i P_s(i) = \sum_{i=1}^{\infty} [i(N-1)+M]\binom{M+i-2}{i-1} P_D^{i-1}(1-P_D)^M \tag{6}$$

The throughput of the SW1 protocol is

$$T = \frac{kM}{n_s n} \tag{7}$$

In the following a code C of type (1023, 983) is considered.

The code C is assumed to be a perfect error-detecting code, i.e. one able to detect all error patterns. This hypothesis allows determination of a lower bound on the throughput. The communication channel has a bit error probability p; therefore, $P_D = 1-(1-p)^n$.

Fig. 2 shows the throughput T of the SW1 protocol against the channel error probability for different values of M for $N = 50$. The throughputs of the classical SW, GBN and SR schemes are also shown for comparison. The gains in throughput using SW1 with respect to the classical SW scheme are quite high for low and medium error probabilities p and with a suitable choice of M. It can be noted that, for mean and high error probabilities, the SW1 scheme presents throughput values near to those of the GBN protocol. However, the SW1 scheme requires a lower energy for information symbol than the GBN scheme.

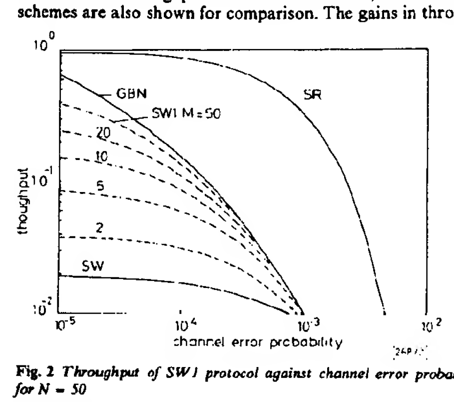

Fig. 2 *Throughput of SW1 protocol against channel error probability for N = 50*

3rd March 1992

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NOVEL FAST GPS/GLONASS CODE-ACQUISITION TECHNIQUE USING LOW UPDATE RATE FFT

A. J. R. M. Coenen and D. J. R. Van Nee

Indexing terms: Radiocommunication, Fourier transforms, Radionavigation

A novel 'differential' decoding technique is proposed which enables pre-averaging instead of postintegration for a substantially low update rate 'FFT-IFT' correlation in spread-spectrum (navigational) receivers. N-channel code acquisition can be performed to monitor the time dispersion with FFT time left to analyse frequency dispersion in highly reflective areas (e.g. an urban environment).

Introduction: Both frequently occurring line-of-sight (LOS) interruptions of NAVSTAR/GPS (the American military global positioning system) or GLONASS (the Russian GPS counterpart) signals and the multipath (MP) signals in the urban environment require a new generation of robust receivers (RXs). MP causes 'time dispersion' of the triangular correlation peak degrading the positioning accuracy. Moreover because of the relative movement between satellite and RX, both LOS and MP signals are affected by the Doppler effect. Especially when an RX is moving at a high speed (e.g. 70 km/h), the carrier Doppler spectrum may show 'frequency dispersion' up to 100 Hz causing false carrier locks and degrading data recovery. With respect to Reference 1, in this Letter an enhanced fast code-acquisition method is proposed to reduce the FFT (fast Fourier transform) and IFT (inverse FFT) update rate (= 1 ms per channel) substantially (> 100 times) to liberate enough FFT time to manage N channels and to calculate an extra N carrier Doppler spectra for frequency dispersion analysis to achieve (not shown here) enhanced data recovery for coherent RX operations by methods such as 'maximal ratio combining' or 'resolvable paths' [2]. The proposed method has led to a Dutch patent application [3] which covers the direct-sequence (DS) spread-spectrum (SS) communication area. It involves a grading in a fast, coarse to finer, code acquisition in a noncoherent way.

Differential decoding: In every single received DSSS LOS signal the code $c(t)$ in the baseband signal $bb(t)$ will be inverted permanently by its data signal $d(t)$ and by its (after down conversion) residual carrier signal $ca(t)$

$$bb(t) = ca(t)d(t)c(t) \tag{1}$$

where $ca(t) = A\cos(2\pi f_{rc}t + \phi)$. By selecting the residual by Doppler ($-5\,\text{kHz} < f_{caDo} < 5\,\text{kHz}$) modulated carrier frequency f_{rc} (e.g. ~ 0 Hz), the carrier evoked inversion rate can be kept relatively low. Compared with the chip rate $f_c = 1/T_c = 1{\cdot}023$ Mchip/s (GPS) or 511 kchip/s (GLONASS) the data inversion rate ($f_d - 1/T_d = 50$ bit/s) is already significantly low. Then, almost all the transition positions in $bb(t)$ are due to code (or chip) edges. The novel proposed correlation technique is based on chip-edge correlation (or matching) regardless of the sign of the edge. Similarly the DPSK (differential phase-shift keying) demodulation [4] can be performed by a 'T_c delay and multiply' operation (i.e. $bb(t) \cdot bb(t - T_c)$). To perform correlation the local generated replica code c_{ref} has to undergo the same differential decoding (DD), i.e. $c_{ref}(t) \cdot c_{ref}(t - T_c)$. However, in practice the SS RX baseband signal-to-noise ratio (SNR) is below zero (GPS, GLONASS: $SNR < -15$ dB). As a penalty for the gained Doppler and data insensitivity, a single differential stage causes a squared increase of the baseband noise power ($SNR < -30$ dB). A partial regain of lost SNR arises by extending this method to M different stages of 'm times T_c delay and multiply' and summing the correlation functions for each M

$$\Delta(SNR_M)_{max} = 10\log M \qquad \text{(dB)} \quad (2)$$

Theoretically, if f_{rc} is adjusted to 0 Hz, eqn. 2 is valid for every M value with $M_{max} = 1023$ (GPS) or 511 (GLONASS). However, as a practical value, $M < 32$ (i.e. $m = \{1, \ldots, 32\}$) should be chosen if $(f_{rc})_{max} < 20$ kHz; then, $32T_c \ll (1/4)(1/f_{rc})$ should be obeyed for an effective DD operation. In that case, every received satellite signal can be processed simultaneously providing the correlator input signal for coarse acquisition.

Pre-averaging by comb filtering: As a second and more profitable recovery method, using the code sequence periodicity $T_p = 1023T_c$ (GPS) or $511T_c$ (GLONASS), a simple integrate and dump (I&D) type comb filter with a feedback delay of T_p to perform an averaging operation enables at least a recovery of the original baseband SNR or even more.

$$\Delta(SNR_{av})_{max} = 10\log b \qquad \text{(dB)} \quad (3)$$

where $b = T_{av}/T_p$ and T_{av} is the averaging time. To avoid time dispersion (caused by a code Doppler f_{coDo} of a few Hertz) of say xT_c ($x < 1$) by averaging, $f_{rc} = f + nf_c/2$ with $n = \{0, 1, 2, \ldots\}$ should be chosen accordingly:

$$f < rx/T_{av} \qquad (4)$$

Here r represents the ratio between the received carrier and chip frequency ($r = f_{caDo}/f_{coDo}$) with $r = 1540$ (GPS), $r \simeq 3000$ (GLONASS). Example: $T_{av} = 0{\cdot}1$ s, $x = 0{\cdot}01$ and $r = 1540$ then $f < 154$ Hz, while eqn. 3 shows a $+20$ dB regain. By averaging, combined with M extended DDs, the recovered SNR according to eqns. 2 and 3 is just enough ($SNR > -13$ dB) to obtain correlation with a detection probability of 0·99 [1]. Here, the process gain by an FFT correlation over 1023 chips amounted to (10 log 1023 =) 30 dB. With respect to Reference 1, choosing $T_{av} > 0{\cdot}1$ s, the FFT update rate can be reduced by more than ($b =$) 100.

Extended differential decoder (EDD): Without efficient use of pseudorandom noise (PRN) properties such as 'cycle and add' (C&A), parameter M and the hardware complexity have a proportional relation. Applying the C&A property coherent to the DD operations on GPS Gold codes G_i ($i = \{1, 2, \ldots, 1023\}$), it can be proven that one G_i changes into two identical Gold-codes (i.e. $m = \{341, 682\}$) plus four groups of 255 other unique Gold codes G_{im} within the class of G_i with 1022 unique code offsets CO_{im} ($m = \{1, 2, \ldots, 1022\}$). For GLONASS, only the code offset of the single linear maximum length sequence (LMLS) of 511 chips will be affected by DD, not the LMLS itself. A proposed structure is shown in Fig. 1 for a GPS system. In Fig. 1 the tapped delay line (TDL) compensates for code offsets for each m value which allows the same code to be originated. Each tap distance per differential stage equals an $m \cdot T_c$ delay. For GPS there are unfortunately only groups of four m values resulting in the same code (e.g. $m = \{8, 25, (1023\text{-}8), 1023\text{-}25\}$), but luckily for GLONASS there is no limitation (e.g. $m = \{1, 2, \ldots, 32\}$). The digitised $bb(t)$ as TDL input signal can be an oversampled two-level signal for ease of multiplication by EXORs. A bit-count operation is followed by an I&D filter as decimator (to $2f_c$ Hz), which forms the input signal to an averager. After T_{av} s the averager register is

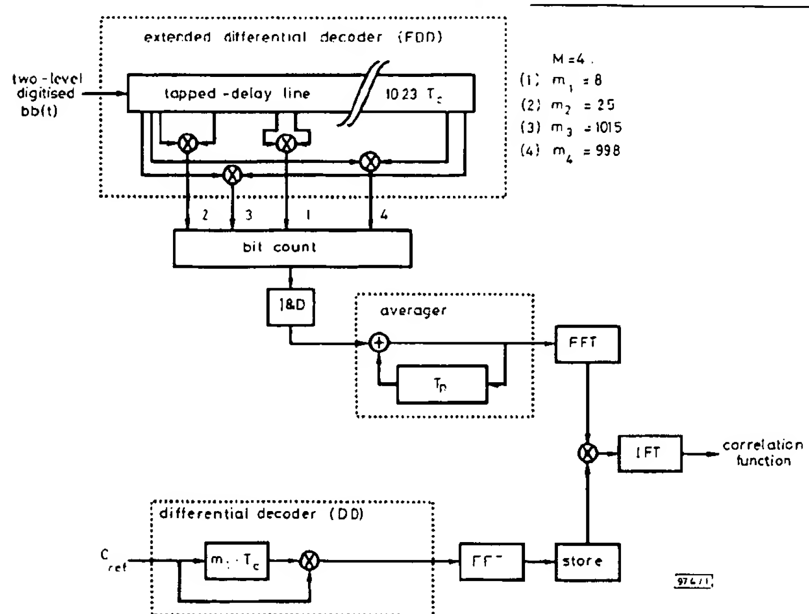

Fig. 1 *Impression of possible structure with EDD/averager combination for GPS*

For example $m = \{8, 25, 998, 1015\}$ leads to same Gold code

EDD for GLONASS can be made much more efficient by its LMLS code property

read out (dumped) to perform correlation (by multiplication in the frequency domain) with c_{ref}. A further increase of M for GPS arises by additionally selecting other groups of four m values. The results should be correlated separately by spectral multiplication, then the product spectra per group are added and finally only one IFFT for the final correlation function generation is applied. According to a strict scheme (not shown here) the TDL in Fig. 1 can be adapted easily for GLONASS for $M = 32$ and $m = (1, 2, \ldots, 32)$. In a cold start situation we may start with $T_{av} = 0{\cdot}1$ s, low m values and $|f_{rc}| < 5$ kHz to find correlation with each of the satellite constellation expected c_{ref} signals coarsely, using only one EDD/averager combination. After a better carrier estimation according to eqn. 4 with one combination per c_{ref}, the code acquisition will be far more precise. Especially in a rate-aided tracking situation T_{av} may increase dramatically.

Conclusions: Differential decoding combined with comb-filter averaging makes the exploitation of the FFT highly efficient for N-channel correlation operations and for frequency-dispersion monitoring. Coarse code acquisition can be performed within the whole Doppler range for a <1 s cold start. Fine control arises by a more precise carrier estimation especially in rate-aided tracking.

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EXTENDED TRENCH-GATE POWER UMOSFET STRUCTURE WITH ULTRALOW SPECIFIC ON-RESISTANCE

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Indexing terms: Field-effect transistors, Transistors, Semiconductor devices and materials

An ultralow specific on-resistance power UMOSFET structure with the trench-gate extending down to the N^- substrate is presented. Specific on-resistances in the range 100–200 $\mu\Omega$ cm^2 have been experimentally demonstrated for devices capable of supporting up to 25 V. Comparison of theoretical and experimental results is provided.

Introduction: Power MOSFETs with breakdown voltages in the 10–50 V range, operated as synchronous rectifiers, have been suggested to replace Schottky diodes in the output stage of power supplies for output voltages below 5 V [1]. For these power MOSFETs to be used as low-voltage rectifiers, however, the power loss in the forward conduction state must be as low as possible, i.e. a further reduction in the device specific on-resistance is demanded. The UMOSFET structure [2–4] is believed to be a good solution for obtaining low specific on-resistance. For 30 V devices, a specific on-resistance of 1000 $\mu\Omega$ cm^2 has been reported by Chang *et al.* [4]. In comparison, 50 V devices have been reported most recently by Matsumoto *et al.* [5] with $R_{on,sp} = 580\,\mu\Omega\,\text{cm}^2$.

This Letter describes a new power MOSFET structure, called the modified mode field-effect transistor (MMFET), having an ultralow on-resistance approaching 100 $\mu\Omega$ cm^2 for a device with a breakdown voltage of 25 V. This improved performance results not only from the inherent features of the UMOSFET structure in eradicating the JFET pinching effect and increasing the channel density, but a unique feature of the MMFET where the current flows from the base to the drain via a highly conductive accumulation layer rather than by spreading into a drift region as in previous UMOSFET structures.

Device structure: The vertical cross-sections of the proposed device (MMFET) and the conventional UMOSFET are shown in Fig. 1. The principal difference between these structures is that in the MMFET, the gate extends into the N^+ substrate. Consequently, the on-state current flows primarily along an accumulation layer formed on the trench sidewall, and, unlike the conventional UMOSFET, the drift region resistance does not contribute to the on-resistance in the MMFET. As a result, the doping concentration can be as low as possible for the drift region with a typical value of 10^{14} cm^{-3}. With such low drift region doping, the maximum sustainable drain voltage for the MMFET device is determined by the punch-through breakdown voltage of the P-base/N^-/N^- structure as long as the oxide in the gate-drain overlapping region is sufficiently thick to support the voltage. For the conventional UMOSFET, however, an optimum

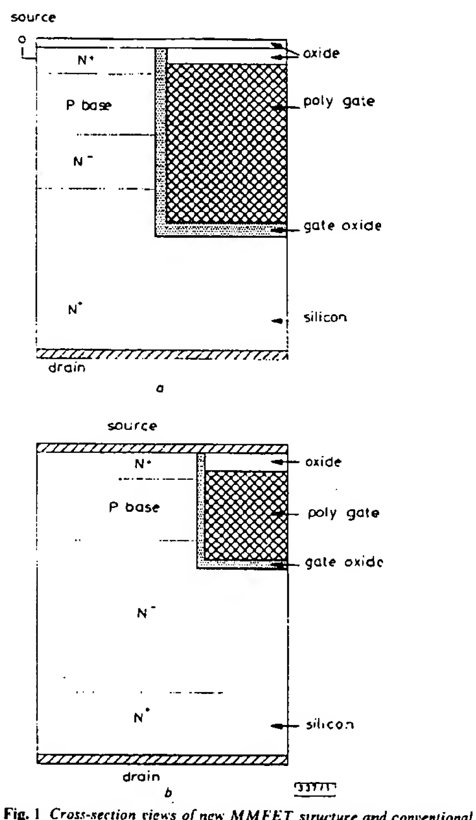

Fig. 1 *Cross-section views of new MMFET structure and conventional UMOSFET structure*
a MMFET
b UMOSFET